# THE ANTENNA LABORATORY

RESEARCH ACTIVITIES in ---

*Automatic Controls* — *Microwave Circuits* — *Terrain Investigations* — *Wave Propagation* — *Antennas* — *Astronautics* — *Radomes* — *Echo Area Studies* — *E M Field Theory* — *Systems Analysis* — *Submillimeter Applications*

FINAL ENGINEERING REPORT
1 March 1959 to 30 April 1960

MICROWAVE CIRCUIT COMPONENTS

Prepared for
Boeing Airplane Company
Seattle 24, Washington

914-4 — 1 May 1960



Department of ELECTRICAL ENGINEERING

THE OHIO STATE UNIVERSITY
RESEARCH FOUNDATION
Columbus 8, Ohio

# REPORT

by

THE OHIO STATE UNIVERSITY RESEARCH FOUNDATION
COLUMBUS 12, OHIO

Cooperator ........................... Boeing Airplane Company
Seattle Division
P. O. Box 3707
Seattle 24, Washington

Investigation of .................... Microwave Circuit Components

Subject of Report ..................... Final Engineering Report
1 March 1959 to 30 April 1960

Submitted by ............................ Antenna Laboratory
Department of Electrical Engineering

Date ................................................ 1 May 1960

"Prepared for the Antennas and Radomes Unit, Electronic Sciences Section, Beoing Airplane Company."

# ABSTRACT

The investigation of transmission line discontinuities which began under the previous contract was completed during the first half of this contract period. This work has led to two useful techniques. 1) A method of determining the characteristic impedance of uniform transmission lines, and 2) a method of transforming an impedance through an unknown discontinuity.

The second half of the period was devoted to a study of metal loss in transmission lines. Coaxial-line, rectangular, and circular waveguide were considered. The effects on the propagation characteristics of operation at temperatures as high as 1500°C were considered and suggestions on minimizing losses are presented.

# TABLE OF CONTENTS

## TABLE OF CONTENTS

## FINAL REPORT

### I. INTRODUCTION

#### A. Transmission-Line Discontinuities

The first portion of this period was devoted to the completion of the investigation of transmission-line discontinuities. Three technical reports were written describing this work.

1. "A Study of Reflections From Small Step Discontinuities in Coaxial Transmission Lines," by C. H. Boehnker, H. N. Dawirs, R. J. Garbacz, E. R. Sapp, and W. S. Tulloss. (Report 914-1, 17 March 1959).

2. "Determination of the Characteristic Impedances of Uniform Transmission Lines," by H.N. Dawirs and W.S. Tulloss. (Report 914-2, 31 July 1959).

3. "Investigation of Small Discontinuities in Transmission Lines," by H.N. Dawirs and W.S. Tulloss. (Report 914-3, 15 August 1959).

Reports 1 and 3 are devoted to applications and further development of a technique first proposed by H.N. Dawirs in " A Simple Slotted-Line Method of Accurately Measuring Impedances Through an Unknown Transition Section," (Report 762-3, 31 December 1957).

Report 2 presents a practical technique for determining the characteristic impedance of uniform transmission lines.

#### B. Lossy Transmission Lines

It has been the purpose of this research to study the effect of metal losses on the transmission characteristics of various types of transmission lines. This problem is becoming important with the increasing need for microwave circuits that will function in high-temperature environments.

The effect of a finite conductivity on the propagation factor and characteristic impedance of the TEM mode in coaxial transmission lines was studied first, both by theoretical and experimental means. This was followed by a review of the literature on rectangular and circular wave guide with two objectives in mind: 1) To find ways of computing the transmission characteristics in the various modes when there is a finite conductivity, and 2) to get ideas which might lead to a lower

attenuation factor.

C. Coaxial Transmission Line

The fundamental equations for the characteristic impedance and propagation factor of a transmission line are

(1) $$Z_o = \sqrt{\frac{R+j\omega L}{G+j\omega C}} \quad \text{ohms and}$$

(2) $$\gamma = \alpha + j\beta = \sqrt{(R+j\omega L)\ (G+j\omega C)}$$

where
R = series resistance per unit length
L = series inductance per unit length
G = shunt conductance per unit length, and
C = shunt capacitance per unit length.

Low-conductivity metals used in the walls of a coaxial line would increase the magnitude of R. Except at very high frequencies it is generally possible to neglect G in commercially produced low-loss cables or in air-filled lines, hence, it is not unreasonable to neglect it in this instance. Since the objective is to minimize the attenuation factor, it is certain that a low-loss dielectric material will be used. It is possible that evacuation of the line would prove to be the best solution since this would also minimize oxidation of the walls at high temperatures.

Neglecting G, equations (1) and (2) become

(3) $$Z_o = \sqrt{\frac{L}{C}} \left( 1 - j\frac{R}{\omega L} \right)^{\frac{1}{2}} \quad \text{ohms and}$$

(4) $$\gamma = j\omega\sqrt{LC} \left( 1 - j\frac{R}{\omega L} \right)^{\frac{1}{2}} .$$

Expansion of the common factor gives

(5) $$\left( 1 - j\frac{R}{\omega L} \right)^{\frac{1}{2}} = 1 - j\frac{1}{2}\left( \frac{R}{\omega L} \right) + \frac{1}{8}\left( \frac{R}{\omega L} \right)^2 + j\frac{1}{16}\left( \frac{R}{\omega L} \right)^3 - \ldots .$$

For ratios of $R/\omega L$ as high as 0.5, all but the first two terms can be omitted. Then,

(6) $$Z_o = \sqrt{\frac{L}{C}} \left( 1 - j\frac{R}{2\omega L} \right) \quad \text{ohms and}$$

(7) $$\gamma = \alpha + j\beta = j\omega\sqrt{LC}\left(1 - j\,\frac{R}{2\omega L}\right) \quad \text{or}$$

(8) $$\alpha = \sqrt{\frac{C}{L}}\,\frac{R}{2}\ .$$

Equation (8) can be put into a more useful form if the following substitutions are made:

(9a) $$L = \frac{\mu}{2\pi}\ln\frac{b}{a}\ \frac{\text{henries}}{\text{meter}}\ ,$$ [10]

(9b) $$C = \frac{2\pi\varepsilon}{\ln b/a}\ \frac{\text{farads}}{\text{meter}}$$

where b is the radius of the outer conductor
a is the radius of the inner conductor.

$\mu$ and $\epsilon$ are the permeability and permittivity, respectively, of the dielectric. Also note that by allowing the inner and outer conductors to have different conductivities, and defining $R = \rho/\delta = \sqrt{\omega\mu/2\sigma}$ (where $\rho$ is the resistivity of the conductor, $\sigma = 1/\rho$ and $\delta$ is the skin depth), the resistance per unit length can be expressed as

(9c) $$R = \frac{R_a}{2\pi a} + \frac{R_b}{2\pi b} = \frac{1}{2\pi}\left(\frac{R_a}{a} + \frac{R_b}{b}\right)\ .$$

Substituting Eq. (9) into Eq. (8) gives

(10a) $$\alpha = \left(\frac{R_a}{a} + \frac{R_b}{b}\right)\frac{1}{2\eta\ln b/a}\ \frac{\text{nepers}}{\text{meter}}$$

where

$$\eta = \sqrt{\mu/\epsilon}\ .$$

This expression for the attenuation is now in the form given by Marcuvitz[2] and its limitations should be clear. Making similar substitutions into Eqs. (6) and (7) gives

(10b) $$Z_o = \frac{1}{2\pi}\,\eta\ln\frac{b}{a} - j\,\frac{1}{4\pi\,\omega\sqrt{\mu\epsilon}}\left(\frac{R_a}{a} + \frac{R_b}{b}\right)$$

and

$$\beta = \omega\sqrt{\mu\epsilon} \tag{10c}$$

If Eq. (10a) is differentiated with respect to a and set equal to zero, it can be shown that for minimum attenuation b = 3.6a. Since this expression determines the characteristic impedance its use may not be practical.

As a means of gaining insight into the problems involved, a study was made of RG-21/U cable. This cable has a nichrome inner conductor which gives it a loss of 46 db/100ft at 1 KMC. Knowing the conductivity, the resistance R was found from Eq. (9c). The capacitance was given by the manufacturer and the inductance was found from Eq. (9a). The conductance was assumed to be negligible. The characteristic impedance and the attenuation factor were then computed using Eqs. (6) and (7). The results are shown in Table I along with the measured insertion loss of a short length of cable.

TABLE I

| | Zo (ohms) | $\alpha$ db/100ft |
|---|---|---|
| Manufacturer's Specifications | 53 | 46.0 |
| Computed Values (1KMC) | 51.8-j0.26 | 41.1 |
| Measured Loss | | 51.0 |

$$R = 50.9 \times 10^{-5}\sqrt{f}\ \frac{\text{ohms}}{\text{meter}},\quad C = 95 \times 10^{-12}\ \frac{\text{farads}}{\text{meter}},$$

$$L = 2.54 \times 10^{-7}\ \frac{\text{henries}}{\text{meter}}$$

It has been pointed out by Moreno[11] that the attenuation factor of flexible cable can be expected to differ from the computed value due to the "braid factor," aging, temperature cycling, flexing, or even nonuniformities introduced at the time of manufacture.

The important thing to note from Table I is the very small imaginary term in the characteristic impedance. (It can also be seen from Eq. (1) that if $G \neq 0$ the imaginary term will be even smaller). Using the dimensions of RG-21/U, both the real and imaginary terms of the characteristic impedance have been plotted in Fig. 1 as a function of resistivity. The resistivities at 1500°C of molybdenum and tungsten, two metals capable of withstanding high temperatures, are marked on the figure to give an indication of the resistivities which are apt to be encountered. In Fig. 2 the attenuation factor is plotted as a function of resistivity. It is, therefore, apparent that in coaxial lines the

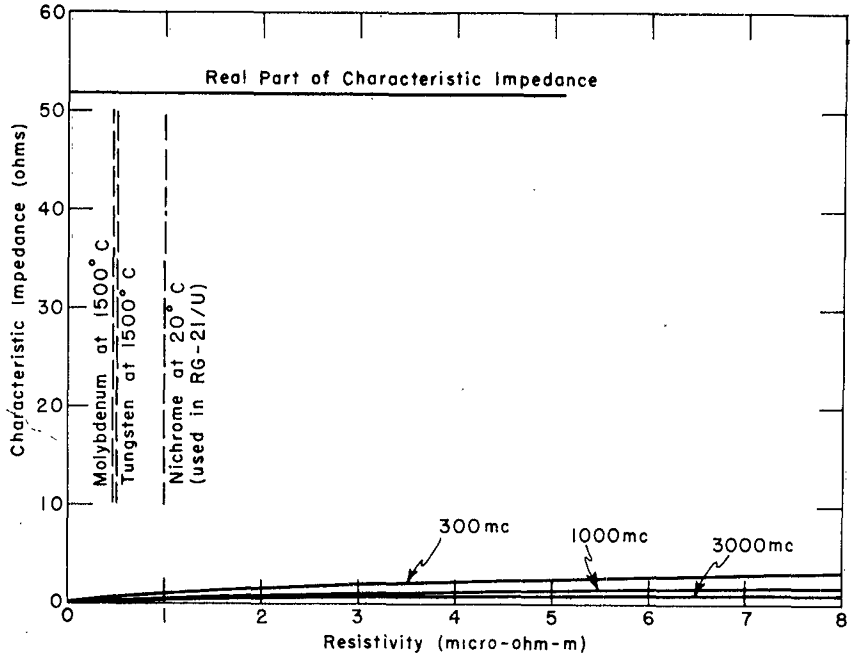


Fig. 1. Characteristic impedance. A comparison of real and imaginary terms.

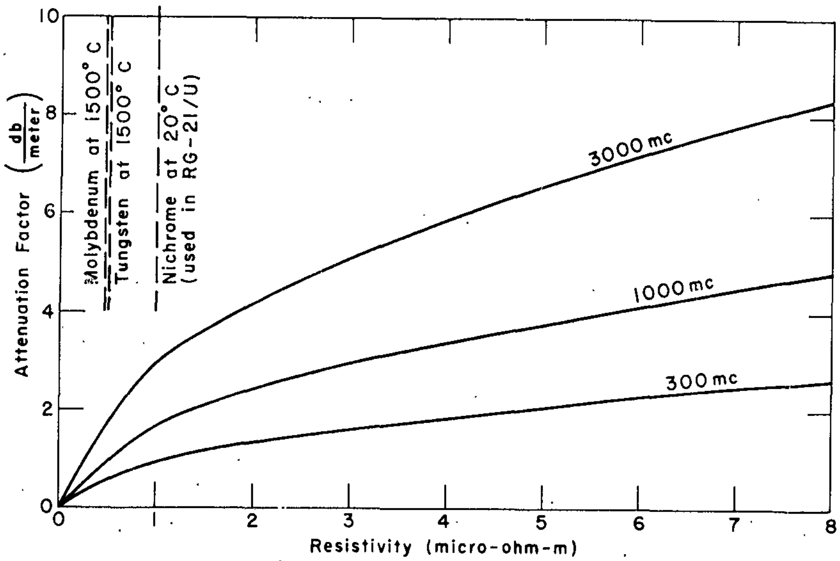


Fig. 2. Attenuation factor

characteristic impedance can be considered to be a real number even though the attenuation factor is appreciable.

Several attempts were made to experimentally determine the propagation factor and characteristic impedance of RG-21/U by means of the equations

$$(11a) \qquad Z_o = \sqrt{Z_{open} \quad Z_{short}} \qquad \text{and}$$

$$(11b) \qquad \tan h\, \gamma \ell = \sqrt{Z_{short} / Z_{open}} ,$$

Where $Z_{short}$ is the input impedance to a length of line terminated in a short circuit, and $Z_{open}$ is the input impedance to the same line terminated by an open circuit. Although the plots of $Z_{open}$ and $Z_{short}$ were self-consistant, indicating good experimental data, the resultant values of $|Z_o|$ varied by as much as 100 percent from what they should have been, in the frequency range covered. It was noticed that most of the error was in $Z_{open}$, due probably to the imperfect open circuit. The discrepancy found in $Z_{short}$ could have been caused by an error of 4 mc in the calibration of the oscillator. Since a calibration error this great does not seem likely, the adapter discontinuities must also have been important.

Apparently the determination of transmission-line characteristics by the open-and short-circuit method is not practical at very high frequencies. It is always possible, however, to measure the attenuation factor by the insertion loss method with considerable accuracy.

## D. Rectangular Waveguide[1, 2, 3, 4, 5, 6, 7]

In the rectangular waveguide shown in Fig. 3, the phase factor is not a linear function of frequency as it is in coaxial lines.

$$(12) \qquad \beta = \omega\sqrt{\epsilon\mu} \quad \sqrt{1 - (\lambda/\lambda_c)^2} \quad \frac{\text{radians}}{\text{meter}}$$

where

$$(13) \qquad \lambda_c = \frac{2ab}{\sqrt{m^2 b^2 + n^2 a^2}}$$

is the cutoff wavelength and, $\lambda = 1/f\sqrt{\mu\epsilon}$ is the wavelength of the propagated wave. The attenuation due to wall loss in a $TE_{mn}$ mode is

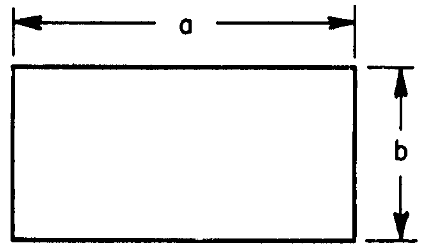


Fig. 3. Waveguide cross-section.

$$(14) \qquad \alpha = \frac{R}{\eta b}\left[ \frac{\epsilon_n m^2 \frac{b}{a} + \epsilon_m n^2}{m^2 \frac{b}{a} + n^2 \frac{a}{b}} \sqrt{1-(\lambda/\lambda_c)^2} + \frac{(\epsilon_n + \epsilon_m \frac{b}{a})(\frac{\lambda}{\lambda_c})^2}{\sqrt{1-(\frac{\lambda}{\lambda_c})^2}} \right] \frac{\text{nepers}}{\text{meter}}$$

where

$\epsilon_i = 1$ if $i = 0$

$\epsilon_i = 2$ if $i \neq 0$

and $\eta = \sqrt{\mu/\epsilon}$ , $R = \sqrt{\frac{\omega\mu}{2\sigma}}$ as before.

For the $TM_{mn}$ modes the attenuation due to wall loss is

$$(15) \qquad \alpha = \frac{2R}{\eta a}\left[ \frac{m^2 + n^2 (\frac{a}{b})^3}{m^2 + n^2 (\frac{a}{b})^2} \right] \left[ \frac{1}{\sqrt{1-(\lambda/\lambda_c)^2}} \right] \frac{\text{nepers}}{\text{meter}}$$

For the dominant mode, $TE_{10}$ , Eq. (14) simplifies considerably to

$$(16) \qquad \alpha_{10} = S_1 + S_2 = \frac{R}{\eta b}\left[ \sqrt{1-(\lambda/2a)^2} + \frac{(1+\frac{2b}{a})(\frac{\lambda}{2a})^2}{\sqrt{1-(\lambda/2a)^2}} \right]$$

The first term, $S_1$, is the loss in the top and bottom planes while the second term comes from the two side walls. Figure 4 shows the relative magnitude of the two terms. These curves suggest ways of decreasing the attenuation in rectangular waveguide.

Doubling b decreases $S_1$, by 50 percent and $S_2$ by 26 percent if a is unchanged. Increasing a raises $S_1$ slightly and lowers $S_2$. It is

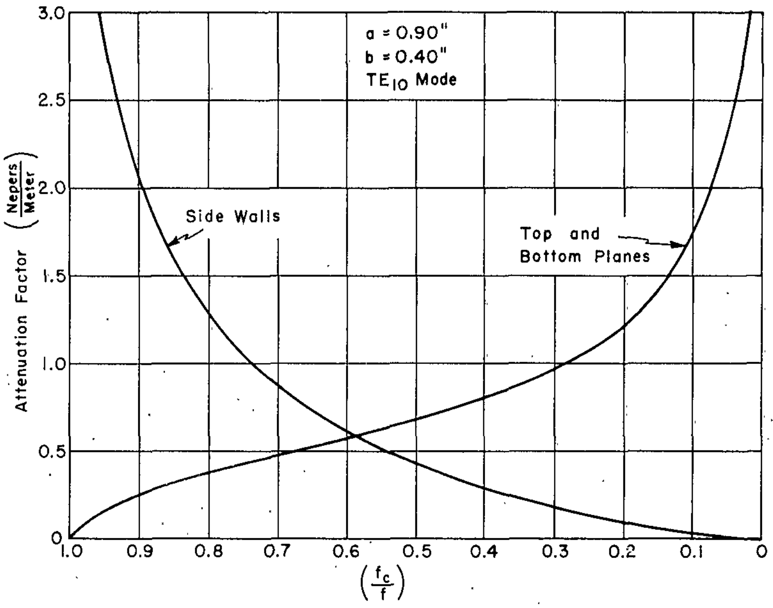


Fig. 4. Losses in rectangular waveguides.

not surprising that an increase in the guide dimensions should decrease the attenuation factor since if both dimensions were made infinite there would be no wall losses. This method of reducing attenuation, however, introduces the problem of mode isolation.

Another interesting possibility is the removal of one set of walls; that is, allow one dimension to become infinite without disturbing the other. Consider first the removal of the top and bottom planes. The loss due to the side walls is very high near cutoff but it decreases to zero as frequency increases. While these loss characteristics could be very useful, they cannot be exploited until a way is found to contain the energy.

One solution to the problem of energy containment for this case has been proposed by F. J. Tischer[12,13,14,15] in what he calls H-guide. Figure 5a shows a cross-sectional view of H-guide. Since much of the energy flows in the dielectric there will, of course, be dielectric losses which are proportional to frequency. However, the total loss for H-guide is considerably less than for rectangular guides[12]. Some modifications of the original form have been successfully tried in an attempt to reduce loss even more. Figure 5b shows one such configuration.

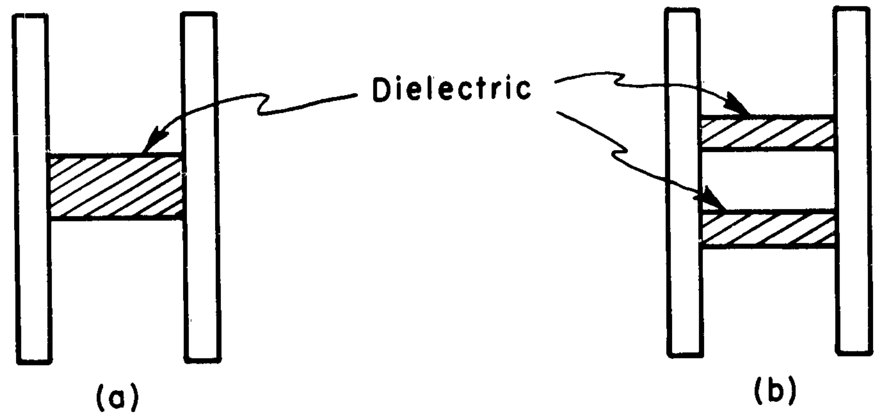


Fig. 5. H-guide cross-sections.

There might also be an advantage in the removal of just the side walls of a rectangular waveguide and operating relatively near the cutoff frequency where the losses are low. Again the line would have to be loaded with dielectric. This guide would have the same cross section as H-guide, the only difference being the orientation of the field components.

Figure 6 shows the attenuation factor of several modes in a rectangular waveguide as a function of increasing frequency or decreasing cutoff frequency where the width-to-height ratio is constant.

Because voltage and current are not clearly defined in waveguide there are several possible definitions of characteristic impedance. The most common definition is that given by Marcuvitz[2]:

$$(17) \qquad Z_o = \frac{E_y}{H_z} = \frac{j\omega\mu}{\gamma}$$

for $TE_{mn}$ modes and

$$(18) \qquad Z_o = \frac{E_y}{H_z} = \frac{\gamma}{j\omega\epsilon}$$

for $TM_{mn}$ modes. Expanding the propagation constant, $\gamma$, gives

$$(19a) \qquad Z_o = \frac{j\omega\mu}{\alpha + j\beta} = \left(\frac{\omega\mu\beta}{\alpha^2 + \beta^2}\right) + j\left(\frac{\omega\mu\alpha}{\alpha^2 + \beta^2}\right)$$

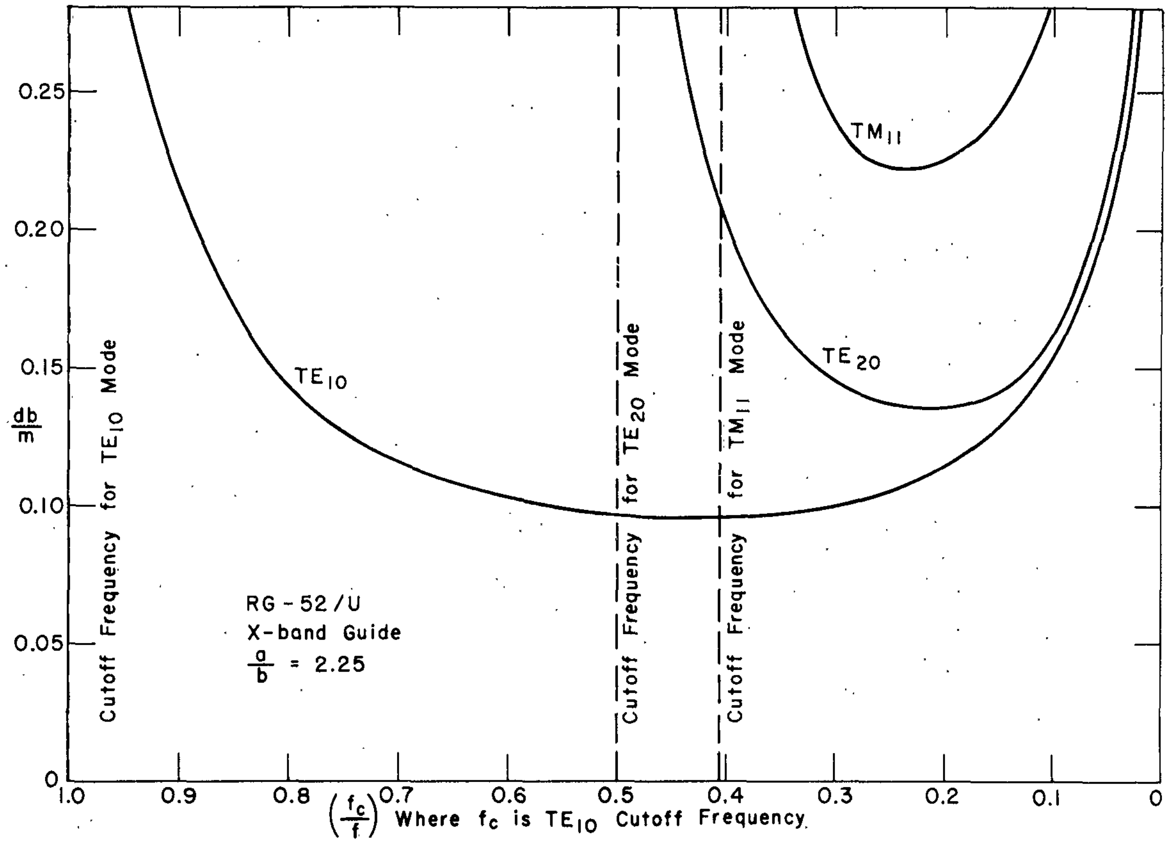


Fig. 6. Attenuation factor in copper rectangular waveguide.

or

$$(19b) \qquad Z_o \simeq \frac{\omega\mu}{\beta} + j\,\frac{\omega\mu\,\alpha}{\beta^2}$$

for $TE_{mn}$ modes , and

$$(20) \qquad Z_o = \frac{\alpha + j\beta}{j\omega\epsilon} = \left(\frac{\beta}{\omega\,\epsilon}\right) - j\left(\frac{\alpha}{\omega\epsilon}\right)$$

for $TM_{mn}$ modes.

To give an indication of the magnitudes involved here, two examples will be presented.

First consider RG-52/U, an X-band waveguide, at a frequency of 10 kmc. For the $\mathbf{TE_{10}}$ mode the attenuation factor is $\alpha = 94.5/\sqrt{\sigma}$ where the conductivity, $\sigma$, is expressed in mhos per meter. Table II shows the propagation factor and the characteristic impedance for three different materials.

TABLE II

| Metal | $\alpha\left(\frac{db}{ft}\right)$ | $\gamma$ | $Z_o$ |
|---|---|---|---|
| Brass at $20^oC$ | 0.0753 | 0.0284+j158 | 500+j0.09 |
| Molybdenum at $1500^oC$ | 1.69 | 0.638+j158 | 500+j2.02 |
| Tungsten at $1500^oC$ | 1.75 | 0.638+j158 | 500+j2.08 |

As a comparison consider RG-91/U, a KU-band guide operating at 15 kmc. It's attenuation factor also in the dominent mode is $\alpha = 148.2/\sqrt{\sigma}$ nepers/meter.

TABLE III

| Metal | $\alpha\left(\frac{db}{ft}\right)$ | $\gamma$ | $Z_o$ |
|---|---|---|---|
| Brass at $20^oC$ | 0.118 | 0.045+j242 | 489+j0.09 |
| Molybdenum at $1500^oC$ | 2.65 | 1.00+j242 | 489+j2.02 |
| Tungsten at $1500^oC$ | 2.76 | 1.04+j242 | 489+j2.10 |

## E. Circular Waveguide[1,5,6]

Circular waveguides have been unpopular for two reasons, primarily; 1) They are narrowband, and 2) the polarization can change as the wave travels down the line.

The attenuation due to wall loss in a $TM_{ne}$ mode is

$$\alpha = \frac{R}{a\eta} \frac{1}{\sqrt{1-(f_c/f)^2}} \quad \frac{\text{nepers}}{\text{meter}} \tag{21}$$

where a is the guide radius and $f_c = P_{ne}/2\pi a\sqrt{\mu\epsilon}$ is the cutoff frequency. The phase constant is given by

$$\beta_{ne} = \sqrt{\omega^2 \mu\epsilon - \left(\frac{P_{ne}}{a}\right)^2} \quad \frac{\text{radian}}{\text{meter}} \tag{22}$$

where $P_{ne}$ is the $\ell$th root of $J_n(x)=0$. For a $TE_{ne}$ mode the attenuation due to wall loss is

$$\alpha = \frac{R}{a\eta} \frac{1}{\sqrt{1-(f_c/f)^2}} \left[\left(\frac{f_c}{f}\right)^2 + \frac{n^2}{(P'_{ne})^2 - n^2}\right] \quad \frac{\text{neper}}{\text{meter}} \tag{23}$$

where $f_c = (P'_{ne})/2\pi a\sqrt{\mu\epsilon}$ and $P'_{ne}$ is the $\ell$th root of $J'_n(x)=0$.

The phase factor differs from that given for the $TM_{ne}$ modes.

$$\beta_{ne} = \sqrt{\omega^2\mu\epsilon - \left(\frac{P'_{ne}}{a}\right)^2} \tag{24}$$

It was pointed out by Chu[1] in 1938 that any wave in a circular guide will have a smaller attenuation factor than does the corresponding wave in a square guide of the same periphery. Figure 7 shows the ratio of the dominant mode attenuation in rectangular guide to that in circular guide where the cutoff frequency is the same for both. The advantage of circular guide from the point of view of loss is quite apparent. Equally apparent in the figure is one of circular guide's big disadvantages-narrow bandwidth. However, by properly exciting the guide and by using mode filters the bandwidth consideration becomes less important. Mode filters do not necessarily add appreciably to the line losses.[8] From both Eqs. (21) and (23) it is apparent that the attenuation factor is inversely proportional to the radius of the guide. But as in rectangular guide, increasing the cross-sectional area decreases $f_c$ and the ratio $(f_c/f)$ if f is kept constant. Figure 8 shows the results of such a change. As

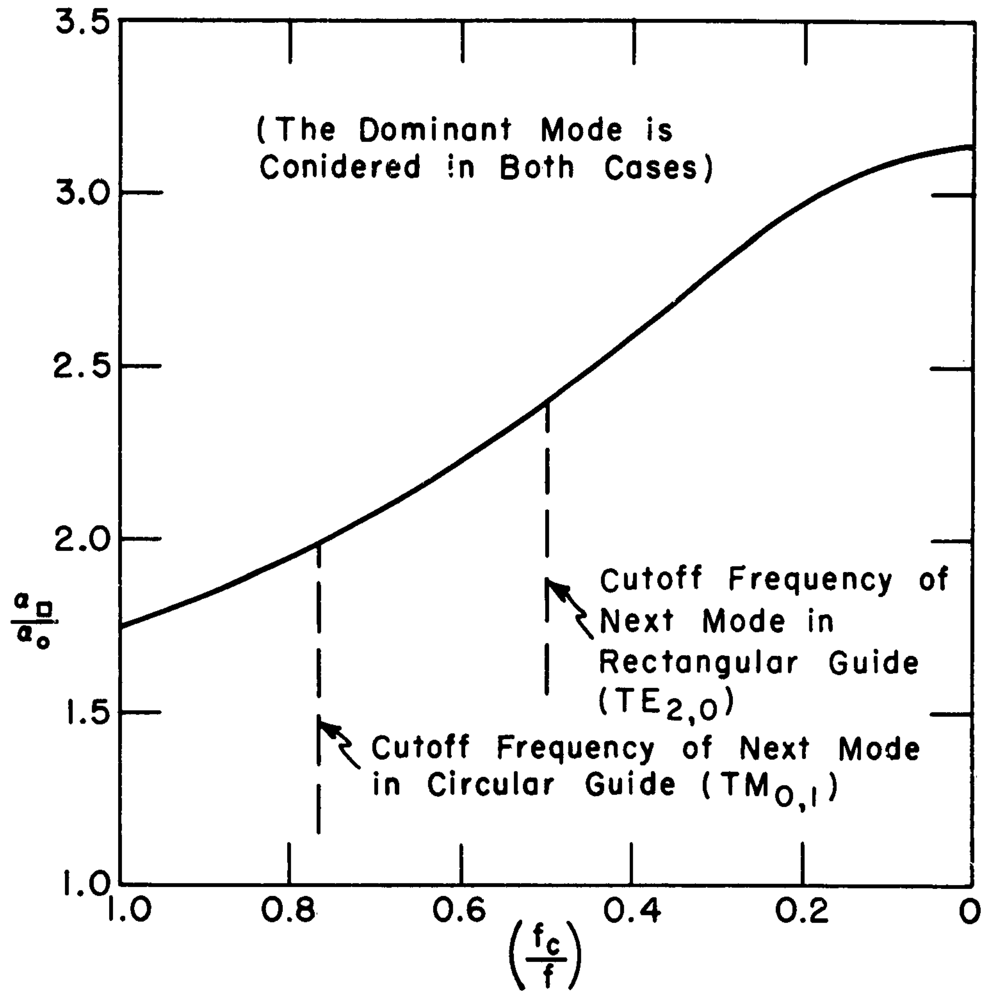


Fig. 7. Ratio of attenuation factor in X-band guide (RG-52/U) to attenuation in circular guide with the same cutoff frequency.

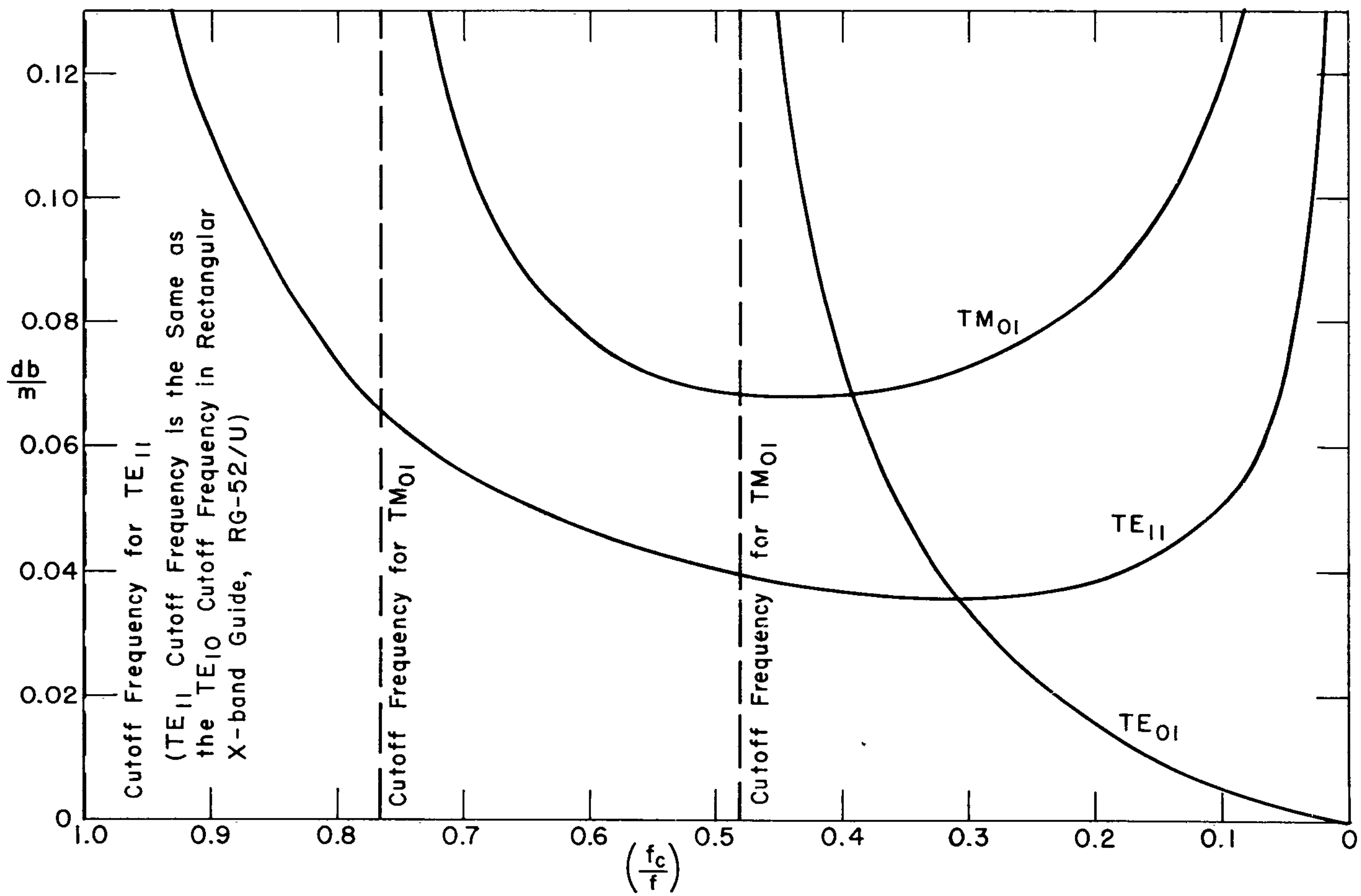


Fig. 8. Attenuation of copper circular waveguide.

( $f_c/f$) becomes smaller than 0.766 it becomes possible for other modes to propagate. From Fig. 8 a decided advantage can be seen in operating with the $TE_{11}$ mode well above cutoff frequency. For all $TM_{n\ell}$ modes the minimum attenuation occurs at $f = \sqrt{3}\, f_c$ where $f_c$ is the cutoff frequency of the mode being considered The $TE_{0\ell}$ modes ( $TE_{01}$ in particular ) appear to be quite promising since attenuation decreases as frequency increases. The attenuation factor for the $TE_{01}$ mode is shown in Fig. 8. So far these modes have not been exploited, chiefly because of their instability. However, this mode has been stuided by Bell Laboratories recently both in solid pipe and helical form as a possible means of long-range transmission of microwaves.[9]

The characteristic impedance of circular waveguide is defined in the same way as for rectangular guide:

$$Z_o = \frac{E_T}{H_T} \cong \frac{\omega\mu}{\beta} + j\,\frac{\omega\mu\,\alpha}{\beta^2} \qquad \text{for } TE_{n\ell} \text{ modes} \tag{25}$$

$$Z_o = \frac{\beta}{\omega\epsilon} - j\,\frac{\alpha}{\omega\epsilon} \qquad \text{for } TM_{n\ell} \text{ modes.} \tag{26}$$

Consider the $TE_{11}$ mode in a circular waveguide which has the same cutoff frequency as the rectangular X-band guide RG-52/U. Again let f = 10 KMC. Table IV shows the propagation constant and characteristic impedance of such a waveguide corresponding to three different materials. The attenuation factor is related to the conductivity ( in mhos/meter) by

$$\alpha = \frac{44.1}{\sqrt{\sigma}} \quad \frac{\text{nepers}}{\text{meter}}$$

TABLE IV

| Metal | $\alpha \left(\frac{db}{ft}\right)$ | $\gamma$ | $Z_o$ |
|---|---|---|---|
| Brass at 20°C | 0.0352 | 0.01328 + j 158 | 499 + j 0.0419 |
| Molybdenum at 1500°C | 0.790 | 0.298 + j 158 | 499 + j 0.941 |
| Tungsten at 1500°C | 0.816 | 0.308 + j 158 | 499 + j 0.973 |

## II. CONCLUSIONS

It was shown that the imaginary term of the characteristic impedance due to metal losses is negligible in coaxial lines, rectangular waveguides, and circular waveguides even for metals with high resistivities. The use of metals which have acceptable structural strength at high temperatures was considered; unfortunately, low conductivities have their greatest effect on the attenuation factor. This effect appears quite serious and indicates a need to redesign present transmission lines to minimize losses or to develop a new type of transmission line.

Some phenomena associated with metal loss in waveguides were pointed out which offer promise of less loss if properly exploited. Several of the suggestions made hinge on the development of relatively lossless mode filters, hence this would seem to be a good starting place in any future research.

## III. BIBLIOGRAPHY


1. Chu, L. J. and Barrow, W. L., "Electromagnetic Waves in Hollow Metal Tubes of Rectangular Cross Section," Proc. I. R. E., Vol. 26, December 1938 p 1520.

2. Marcuvitz, N., Waveguide Handbook, No. 10, Radiation Laboratory Series, McGraw-Hill.

3. Slater, J. C., Microwave Transmission, McGraw-Hill, p. 138.

4. Schelkunoff, S. A., Electromagnetic Waves, D. Van Nostrand Co.

5. Sarbacher, R. I. and Edson, W. A., Hyper and Ultrahigh Frequency Engineering, John Wiley and Sons, Inc., p 214.

6. Ramo, S. and Whinnery, J. R., Fields and Waves in Modern Radio, John Wiley and Sons, Inc., p. 344.

7. Schelkunoff, S. A., "Transmission Theory of Plane Electromagnetic Waves," Proc. I. R. E., 1937, Vol. 25, p. 1457.

8. Warters, W. D., "The Effects of Mode Filters on the Transmission Characteristics of Circular Electric Waves in a Circular Waveguide," Bell Sys. Tech. J., Vol. 37, pp. 657-677, May 1958.

9. Warters, W. D. and Rowe, H. W., "The Effects of Mode Conversion in Long Circular Waveguide," 1958 IRE-WESCON Convention Record, Part 1, p. 13.

10. Karakash, J. J., Transmission Lines and Filter Networks, The Macmillan Company, p. 22.

11. Moreno, T., Microwave Transmission Design Data, Dover Publications Inc., p. 73.

12. Abdel-Bary, A. F. Mohammed, "Multi-Slab Isotropic Dielectric Waveguides," Thesis, The Ohio State University, 1959.

13. Tischer, F. J., "The H-Guide, A Waveguide for Microwaves," IRE Convention Record, 1956, Part 5.

14. Tischer, F. J., "Properties of the H-Guide at Microwaves and Millimeter Wave," 1958 IRE-WESCON Convention Record, Part I.

15. Wu, Chen Pang, "A Study of the H-Guide Attenuation Due to Non-Conventional Influences Such as Radiation, Discontinuities, Non-Homogeneities and Excitation of Spurious Modes," Thesis, The Ohio State University, 1959.

Investigator Edward K Damon Date 18 Mar 60

Investigator ........ Date ........

Investigator ........ Date ........

Investigator ........ Date ........

Supervisor T E Tice by R G Fouty Date 18 March 60

For The Ohio State University Research Foundation

Executive Director Orvan C. Woolpert LLS Date March 21, 1960